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Chapter 17

INDUCTION MACHINE DESIGN FOR VARIABLE SPEED

17.1 INTRODUCTION

Variable speed drives with induction motors are by now a mature technology with strong and dynamic markets for applications in all industries. Based on the load torque/speed envelope, three main types of applications may be distinguished:

- Servodrives: no constant power speed range
- General drives: moderate constant power speed range ($\omega_{\max}/\omega_b \leq 2$)
- Constant power drives: large constant power speed range ($\omega_{\max}/\omega_b \geq 2$)

Servodrives for robots, machine tools, are characterized, in general, by constant torque versus speed up to base speed ω_b . The base speed ω_b is the speed for which the motor can produce (for continuous service) for rated voltage and rated temperature rise, the rated (base) power P_b , and rated torque T_{eb} .

Servodrives are characterized by fast torque and speed response and thus for short time, during transients, the motor has to provide a much higher torque T_{ek} than T_{eb} . The higher the better for speed response quickness. Also servodrives are characterized by sustained very low speed and up to rated (base) torque operation, for speed or position control. In such conditions, low torque pulsations and limited temperature rise are imperative.

Temperature rise has to be limited to avoid both winding insulation failure and mechanical deformation of the shaft which would introduce errors in position control.

In general, servodrives have a constant speed (separate shaft) power, grid fed, ventilator attached to the IM at the non-driving end. The finned stator frame is thus axially cooled through the ventilator's action. Alternatively, liquid cooling of the stator may be provided.

Even from such a brief introduction, it becomes clear that the design performance indexes of IMs for servodrives need special treatment. However, fast torque and speed response and low torque pulsations are paramount. Efficiency and power factor are second order performance indexes as the inverter KVA rating is designed for the low duration peak torque (speed) transients requirements.

General drives, which cover the bulk of variable speed applications, are represented by fans, pumps, compressors, etc.

General drives are characterized by a limited speed control range, in general, from $0.1\omega_b$ to $2\omega_b$. Above base speed ω_b constant power is provided. A limited constant power speed range $\omega_{\max}/\omega_b = 2.0$ is sufficient for most cases. Above base speed, the voltage stays constant.

Based on the stator voltage circuit equation at steady state,

$$\bar{V}_s = \bar{I}_s R_s + j\omega_1 \bar{\Psi}_s \quad (17.1)$$

with $R_s \approx 0$

$$\bar{\Psi}_s \approx \frac{\bar{V}_s}{\omega_1} \quad (17.2)$$

Above base speed (ω_b), the frequency ω_1 increases for constant voltage. Consequently, the stator flux level decreases. Flux weakening occurs. We might say that general drives have a 2/1-flux weakening speed range.

As expected, there is some torque reserve for fast acceleration and braking at any speed. About 150% to 200% overloading is typical.

General drives use IMs with on-the-shaft ventilators. More sophisticated radial-axial cooling systems with a second cooling agent in the stator may be used.

General drives may use high efficiency IM designs as in this case efficiency is important.

Made with class F insulated preformed coils and insulated bearings for powers above 100 kW and up to 2000 kW, and at low voltage (maximum 690 V), such motors are used in both constant and variable speed applications. While designing IMs on purpose for general variable speed drives is possible, it may seem more practical to have a single design both for constant and variable speed: the high efficiency induction motor.

Constant power variable speed applications, such as spindles or hybrid (or electric) car propulsion, generator systems, the main objective is a large flux weakening speed range $\omega_{\max}/\omega_b > 2$, in general more than 3–4, even 6–7 in special cases. Designing an IM for a wide constant power speed range is very challenging because the breakdown torque T_{bk} is in p.u. limited: $t_{bk} < 3$ in general.

$$T_{cK} \approx \frac{3p_1}{2} \left(\frac{V_{ph}}{\omega_1} \right)^2 \cdot \frac{1}{L_{sc}} \quad (17.3)$$

Increasing the breakdown torque as the base speed (frequency) increases could be done by

- Decreasing the pole number $2p_1$
 - Increasing the phase voltage
 - Decreasing the leakage inductance L_{sc} (by increased motor size, winding tapping, phase connection changing, special slot (winding) designs to reduce L_{sc})
- Each of these solutions has impact on both IM and static power converter costs. The global cost of the drive and the capitalized cost of its losses are solid criteria for appropriate designs. Such applications are most challenging. Yet another category of variable speed applications is represented by super-high speed drives.

- For fast machine tools, vacuum pumps etc., speeds which imply fundamental frequencies above 300 Hz (say above 18,000 rpm) are considered here for up to 100 kW powers and above 150 Hz (9000 rpm) for higher powers.

As the peripheral speed goes up, above (60–80) m/s, the mechanical constraints become predominant and thus novel rotor configurations become necessary. Solid rotors with copper bars are among the solutions considered in such applications. Also, as the size of IM increases with torque, high-speed machines tend to be small (in volume/power) and thus heat removal becomes a problem. In many cases forced liquid cooling of the stator is mandatory.

Despite worldwide efforts in the last decade, the design of IMs for variable speed, by analytical and numerical methods, did not crystallize in widely accepted methodologies.

What follows should be considered a small step towards such a daring goal.

As basically the design expressions and algorithms developed for constant V/f (speed) are applicable to variable speed design, we will concentrate only on what distinguishes this latter enterprise.

- In the end, a rather detailed design example is presented. Among the main issues in IM design for variable speed, we treat here
- Power and voltage derating
- Reducing skin effect
- Reducing torque pulsations
- Increasing efficiency
- Approaches to leakage inductance reduction
- Design for wide constant power wide speed range
- Design for variable very high speed

17.2 POWER AND VOLTAGE DERATING

An induction motor is only a part of a variable speed drive assembly (Figure 17.1).

As such, the IM is fed from the power electronics converter (PEC) directly, but indirectly, in most cases, from the industrial power grid.

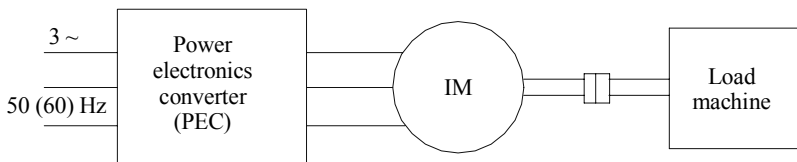


Figure 17.1 Induction machine in a variable speed drive system

There are a few cases where the PEC is fed from a dc source (battery).

The PEC inflicts on the motor voltage harmonics (in a voltage source type) or current harmonics (in a current source type). In this way, voltage and current

harmonics, whose frequency and amplitude are dependent on the PWM (control) strategy and power level, are imposed on the induction motor.

Additionally, high frequency common voltage mode currents may occur in the stator phases in high frequency PWM voltage source converter IM drives. All modern PECs have incorporated filtering methods to reduce the additional current and voltage (flux) harmonics in the IMs as they produce additional losses.

Analytical and even finite element methods have been proposed to cater to these time harmonics losses (see Chapter 11). Still, these additional time harmonics core and winding losses depend not only on machine geometry and materials, but also on PWM, switching frequency and load level. [1,2]

On top of this, for given power grid voltage, the maximum fundamental voltage at motor terminals depends on the type of PEC (single or double stage, type of power electronics switches (PES)) and PWM (control) strategy.

Though each type of PEC has its own harmonics and voltage drop signature, the general rule is that lately both these indexes have decreased. The matrix converter is a notable exception in the sense that its voltage derating (drop) is larger (up to 20%) in general.

Voltage derating – less than 10%, in general 5%–means that the motor design is performed at a rated voltage V_m which is smaller than the a.c. power grid voltage V_g :

$$V_m = V_g (1 - v_{\text{derat}}) ; v_{\text{derat}} < 0.1 \quad (17.4)$$

Power derating comes into play in the design when we choose the value of Esson's constant C_0 (W/m^3), as defined by past experience for sinusoidal power supply, and reduce it to C'_0 for variable V/f supply:

$$C'_0 = C_0 (1 - p_{\text{derat}}) ; p_{\text{derat}} \approx (0.08 - 0.12) \quad (17.5)$$

It may be argued that this way of handling the PEC-supplied IM design is quite empirical. True, but this is done only to initiate the design (sizing) process. After the sizing is finished, the voltage drops in the PEC and the time harmonics core and winding losses may be calculated (see Chapter 11). Then design refinements are done. Alternatively, if prototyping is feasible, test results are used to validate (or correct) the loss computation methodologies.

There are two main cases: one when the motor exists, as designed for sinusoidal power supply, and the other when a new motor is to be designed for the scope.

The derating concepts serve both these cases in the same way.

However, the power derating concept is of little use where no solid past experience exists, such as in wide constant power speed range drives or in super-high speed drives. In such cases, the tangential specific force (N/cm^2), Chapter 14, with limited current sheet (or current density) and flux densities, seem to be the right guidelines for practical solutions. Finally, the temperature rise and performance (constraints) checks may lead to design iterations. As

already mentioned in Chapter 14, the rated (base) tangential specific force (σ_t) for sinusoidal power supply is

$$\sigma_t^{\sin} \approx (0.3 - 4.0) \text{ N / cm}^2 \quad (17.6)$$

Derating now may be applied to $\sigma_t^{\sin}$ to get σ_t^{PEC}

$$\sigma_t^{\text{PEC}} = \sigma_t^{\sin} (1 - p_{\text{derat}}) \quad (17.7)$$

for same rated (base) torque and speed.

The value of σ_t^{PEC} increases with rated (base) torque and decreases with base speed.

17.3 REDUCING THE SKIN EFFECT IN WINDINGS

In variable speed drives, variable V and f are used. Starting torque and current constraints are not relevant in designing the IM. However, for fast torque (speed) response during variable frequency and voltage starting or loading or for constant power wide speed range applications, the breakdown torque has to be large.

Unfortunately, increasing the breakdown torque without enlarging the machine geometry is not an easy task.

On the other hand, rotor skin effect that limits the starting current and produces larger starting torque, based on a larger rotor resistance is no longer necessary.

Reducing skin effect is now mandatory to reduce additional time harmonics winding losses.

Skin effect in winding losses depends on frequency, conductor size, and position in slots. First, the rotor and stator skin effect at fundamental frequency is to be reduced. Second, the rotor and stator skin effect has to be checked and limited at PEC switching frequency. The amplitude of currents is larger for the fundamental than for time harmonics. Still the time harmonics conductor losses at large switching frequencies are notable. In super-high speed IMs the fundamental frequency is already large, (300-3(5)000) Hz. In this case the fundamental frequency skin effects are to be severely checked and kept under control for any practical design as the slip frequency may reach tenth of Hz (up to 50-60 Hz).

As the skin effect tends to be larger in the rotor cage we will start with this problem.

Rotor bar skin effect reduction

The skin effect is a direct function of the parameter:

$$\zeta = h \sqrt{\frac{S \omega_l \sigma_{\text{cor}} \mu_0}{2}} \quad (17.8)$$

The slot shape also counts. But once the slot is rectangular or circular, only the slot diameter, and respectively, the slot height counts.

Rounded trapezoidal slots may also be used to secure constant tooth flux density and further reduce the skin effects (Figure 17.2).

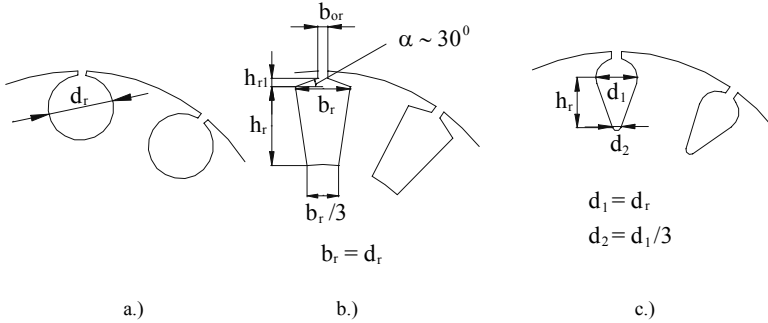


Figure 17.2 Rotor bar slots with low skin effect
a.) round shape b.) rectangular c.) pear-shape

For given rotor slot (bar) area A_b (Figure 17.2 a,b,c), we have

$$\begin{aligned}
 A_b &= \frac{\pi d_r^2}{4} \\
 &= h_r \cdot \frac{2d_r}{3} + \frac{h_{r1}}{2} (b_{or} + d_r) \\
 &= h_r \frac{2}{3} d_1 + \frac{\pi 5}{36} d_r^2
 \end{aligned} \tag{17.9}$$

For the rectangular slot, the skin effect coefficients K_r and K_x have the standard formulas

$$\begin{aligned}
 K_R &= \zeta \frac{\sinh 2\zeta + \sin 2\zeta}{\cosh 2\zeta - \cos 2\zeta} \\
 K_x &= \frac{3}{2\zeta} \frac{\sinh 2\zeta - \sin 2\zeta}{\cosh 2\zeta - \cos 2\zeta}
 \end{aligned} \tag{17.10}$$

In contrast, for round or trapezoidal-round slots, the multiple-layer approach of Chapter 9, has to be used.

A few remarks are in order:

- As expected, for given geometry and slip frequency, skin effects are more important in copper than in aluminum bars
- For given rotor slot area, the round bar has limited use
- As the bar area (bar current or motor torque) increases, the maximum slip frequency $f_r = S f_1$ for which $K_R < 1.1$ diminishes
- Peak slip frequency f_{srk} varies from 2 Hz to 10 Hz
- The smaller values correspond to larger (MW) machines and larger values to subKW machines designed for base frequencies of 50 (60) Hz. For f_{srk} , $K_R < 1.1$ has to be fulfilled if rotor additional losses are to

be limited. Consequently, the maximum slot depth depends heavily on motor peak torque requirements

- For super-high speed machines, f_{srk} may reach even 50 (60) Hz, so extreme care in designing the rotor bars is to be exercised (in the sense of severe limitation of slot depth, if possible)
- Maintaining reduced skin effect at f_{srk} means, apparently, less deep slots and thus, for given stator bore diameter, longer lamination stacks. As shown in the next paragraph this leads to slightly lower leakage inductances, and thus to larger breakdown torque. That is, a beneficial effect.
- When the rotor skin effect for f_{srk} may not be limited by reducing the slot depth, we have to go so far as to suggest the usage of a wound rotor with shortcircuited phases and mechanically enforced end-connections against centrifugal forces
- To reduce the skin effect in the end rings, they should not be placed very close to the laminated stack, though their heat transmission and mechanical resilience is a bit compromised
- Using copper instead of aluminium leads to a notable reduction of rotor bar resistance for same bar cross-section though the skin effect is larger. A smaller copper bar cross-section is allowed, for same resistance as aluminum, but for less deep slots and thus smaller slot leakage inductance. Again, larger breakdown torque may be obtained. The extracost of copper may prove well worth while due to lower losses in the machine.
- As the skin effect is maintained low, the slot-body geometrical specific permeance λ_{sr} for the three cases mentioned earlier (Figure 17.1) is:

$$\begin{aligned}\lambda_{sr}^{round} &\approx 0.666 \\ \lambda_{sr}^{rect} &\approx \frac{h_r}{3b_r} + \frac{2}{3\sqrt{3}} \\ b_r &= d_r \\ \lambda_{sr}^{trap} &\approx \frac{h_r}{2d_r} + 0.4\end{aligned}\tag{17.11}$$

Equations (17.11) suggest that, in order to provide for identical slot geometrical specific permeance λ_{sr} , $h_r/b_r \leq 1.5$ for the rectangular slot and $h_r/d_r < 0.5$ for the trapezoidal slot. As the round part of slot area is not negligible, this might be feasible ($h_r/d_r \approx \pi/8 < 0.5$), especially for low torque machines.

Also for the rectangular slot with $b_r = d_r$, $h_r = (\pi/4) d_r \ll 1.5$, so the rectangular slot may produce

$$\lambda_{sr}^{root}(\pi/4) = \pi/12 + 2/3\sqrt{3} = 0.67 \approx \lambda_{sr}^{round}\tag{17.11'}$$

In reality, as the rated torque gets larger, the round bar is difficult to adopt as it would lead to a too small number of rotor slots or too a larger rotor

diameter. In general, a slot aspect ratio $h_r/b_r \leq 3$ may be considered acceptable for many practical cases.

- The skin effect in the stator windings, at least for fundamental frequencies less than 100(120) Hz is negligible in well designed IMs for all power levels. For large powers, elementary rectangular cross section conductors in parallel are used. They are eventually stranded in the end-connection zone. The skin effect and circulating current additional losses have to be limited in large motors
- In super-high speed IMs, for fundamental frequencies above 300 Hz (up to 3 kHz or more), stator skin effect has to be carefully investigated and suppressed by additional methods such as Litz wire, or even by using thin wall pipe conductors with direct liquid cooling when needed
- Skin-effect stator and rotor winding losses at PWM inverter carrier frequency are to be calculated as shown in Chapter 11, paragraph 11.12.

17.4 TORQUE PULSATIONS REDUCTION

Torque pulsations are produced both by airgap flux density space harmonics in interaction with stator (rotor) m.m.f. space harmonics and by voltage (current) time harmonics produced by the power electronics converter (PEC) which supplies the IM to produce variable speed.

As torque time harmonics pulsations depend mainly on the PEC type and power level we will not treat them here. The space harmonic torque pulsations are produced by the so called parasitic torques (see Chapter 10). They are of two categories: asynchronous and synchronous and depend on the number of rotor and stator slots, slot opening/airgap ratios and airgap/pole pitch ratio, and the degree of saturation of stator (rotor) core. They all however occur at rather large values of slip: $S > 0.7$ in general.

This fact seems to suggest that for pump/fan type applications, where the minimum speed hardly goes below 30% base speed, the parasitic torques occur only during starting.

Even so, they should be considered, and the same rules apply, in choosing stator rotor slot number combinations, as for constant V and f design (Chapter 15, [table 15.5](#)).

- As shown in Chapter 15, slot openings tend to amplify the parasitic synchronous torques for $N_r > N_s$ (N_r – rotor slot count, N_s – stator slot count). Consequently $N_r < N_s$ appears to be a general design rule for variables V and f, even without rotor slot skewing (for series connected stator windings).
- Adequate stator coil throw chording (5/6) will reduce drastically asynchronous parasitic torque.
- Carefully chosen slot openings to mitigate between low parasitic torques and acceptable slot leakage inductances are also essential.
- Parasitic torque reduction is all the more important in servodrive applications with sustained low (even very low) speed operation. In

such cases, additional measures such as skewed resin insulated rotor bars and eventually closed rotor slots and semiclosed stator slots are necessary. FEM investigation of parasitic torques may become necessary to secure sound results.

17.5 INCREASING EFFICIENCY

Increasing efficiency is related to loss reduction. There are fundamental core and winding losses and additional ones due to space and time harmonics.

Lower values of current and flux densities lead to a larger but more efficient motor. This is why high efficiency motors are considered suitable for variables V and f .

Additional core losses and winding losses have been treated in detail in Chapter 11.

Here we only point out that the rules to reduce additional losses, presented in Chapter 11 still hold. They are reproduced here and extended for convenience and further discussion.

- Large number of slots/pole/phase in order to increase the order of the first slot space harmonic
- Insulated or uninsulated high bar-slot wall contact resistance rotor bars in long stack skewed rotors, to reduce interbar current losses
- Skewing is not adequate for low bar-slot wall contact resistance as it does not reduce the harmonics (stray) cage losses while it does increase interbar current losses
- $0.8N_s < N_r < N_s$ – to reduce the differential leakage coefficient of the first slot harmonics ($N_s \pm p_1$), and thus reduce the interbar current losses
- For $N_r < N_s$ skewing may be altogether eliminated after parasitic torque levels are checked. For $q = 1, 2$ skewing seems mandatory
- Usage of thin special laminations (0.1 mm) above $f_{in} = 300\text{Hz}$ is recommended to reduce core loss in super-high speed IM drives
- Chorded coils ($y/\tau \approx 5/6$) reduce the asynchronous parasitic torque produced by the first phase belt harmonic ($v = 5$)
- With delta connection of stator phases: $(N_s - N_r) \neq 2p_1, 4p_1, 8p_1$.
- With parallel paths stator windings, the stator interpath circulating currents produced by rotor bar current m.m.f. harmonics have to be avoided by observing certain symmetry of stator winding paths
- Small stator (rotor) slot openings lead to smaller surface and tooth flux pulsation additional core losses but they tend to increase the leakage inductances and thus reduce the breakdown torque
- Carefully increase the airgap to reduce additional core and cage losses without compromising too much the power factor and efficiency
- Use sharp tools and annealed laminations to reduce surface core losses
- Return core losses rotor surface to prevent rotor lamination shortcircuits which would lead to increased rotor surface core losses

- Use only recommended N_s , N_r combinations and check for parasitic torque and stray load levels
- To reduce the time and space harmonics losses in the rotor cage, U shape bridge rotor slots have been proposed (Figure 17.3). [3]

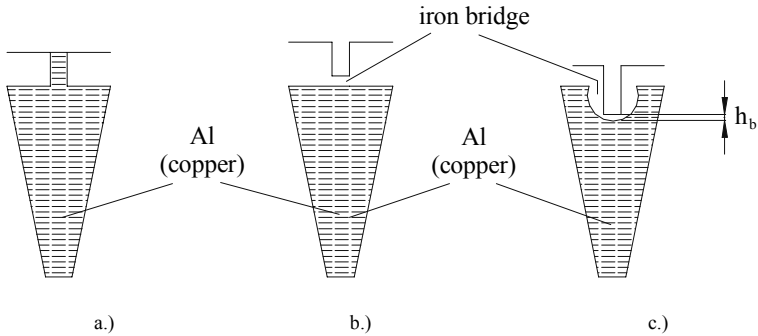


Figure 17.3 Rotor slot designs
a) conventional b) straight bridge closed slot c) u-bridge close slot

In essence in conventional rotor slots, the airgap flux density harmonics induce voltages which produce eddy currents in the aluminium situated in the slot necks. By providing a slit in the rotor laminations (Figure 17.3b, 17.3c), the rotor conductor is moved further away from the airgap and thus the additional cage losses are reduced.

However, this advantage comes with three secondary effects.

First, the eddy currents in the aluminium cage close to airgap damp the airgap flux density variation on the rotor surface and in the rotor tooth. This, in turn, limits the rotor core surface and tooth flux pulsation core losses.

In our new situation, it no longer occurs. Skewed rotor slots seem appropriate to keep the rotor surface and tooth flux pulsation core losses under control.

Second, the iron bridge height h_b above the slot, even when saturated, leads to a notable additional slot leakage geometrical permeance coefficient: λ_b .

Consequently, the value of L_{sc} is slightly increased, leading to a breakdown torque reduction.

Third, the mechanical resilience of the rotor structure is somewhat reduced which might prevent the usage of this solution to super-high speed IMs.

17.6 INCREASING THE BREAKDOWN TORQUE

As already inferred, a large breakdown torque is desirable either for high transient torque reserve or for widening the constant power speed range. Increasing the breakdown torque boils down to **leakage inductance decreasing**, when the base speed and stator voltage are given (17.3).

The total leakage inductance of the IM contains a few terms as shown in Chapter 6.

The stator and rotor leakage inductances are (Chapter 6)

$$L_{sl} = 2\mu_0 L_i \cdot n_s^2 p_1 q_1 (\lambda_{ss} + \lambda_{zs} + \lambda_{ds} + \lambda_{end}) \quad (17.12)$$

n_s – conductors slot

$$L_{rl} = 4m_1 \frac{(W_i K_{w1})^2}{N_r} \times 2\mu_0 L_i (\lambda_b + \lambda_{er} + \lambda_{zr} + \lambda_{dr} + \lambda_{skew}) \quad (17.13)$$

- m_1 – number of phases
- L_i – stack length
- q_1 – slots/pole/phase
- λ_{ss} – stator slot permeance coefficient
- λ_{zs} – stator zig-zag permeance coefficient
- λ_{ds} – stator differential permeance coefficient
- λ_{end} – stator differential permeance coefficient
- λ_b – rotor slot permeance coefficient
- λ_{er} – end ring permeance coefficient
- λ_{zr} – rotor zig-zag permeance coefficient
- λ_{dr} – rotor differential permeance coefficient
- λ_{skew} – rotor skew leakage coefficient

The two general expressions are valid for open and semi-closed slots. For closed rotor slots λ_b has the slot iron bridge term as rotor current dependent.

With so many terms, out of which very few may be neglected in any realistic analysis, it becomes clear that an easy sensitivity analysis of L_{sl} and L_{rl} to various machine geometrical variables is not easy to wage.

The main geometrical variables which influence $L_{sc} = L_{sl} + L_{rl}$ are

- Pole number: $2p_1$
- Stack length/pole pitch ratio: L_i/τ
- Slot/tooth width ratio: b_{s1r}/b_{t1r}
- Stator bore diameter
- Stator slots/pole/phase q_1
- Rotor slots/pole pair N_r/p_1
- Stator (rotor) slot aspect ratio h_{ss1r}/b_{s1r}
- Airgap flux density level B_g
- Stator (rotor) base torque (design) current density

Simplified sensitivity analysis [4] of j_{cos} , j_{Al} to t_{sc} (or t_{bk} – breakdown torque in p.u.) have revealed that 2,4,6 poles are the main pole counts to consider except for very low direct speed drives – conveyor drives – where even $2p_1 = 12$ is to be considered, only to eliminate the mechanical transmission.

Globally, when efficiency, breakdown torque, and motor volume are all considered, the 4 pole motor seems most desirable.

Reducing the number of poles to $2p_1 = 2$, for given speed $n_1 = f_1/p_1$ (rps) means lower frequency, but for the same stator bore diameter, it means larger pole pitch and longer end connections and thus, larger λ_{end} in (17.12).

Now with q_1 larger, for $2p_1 = 2$ the differential leakage coefficient λ_{ds} is reduced. So, unless the stack length is not small at design start (pancake shape), the stator leakage inductance (17.12) decreases by a ratio of between 1 and 2 when the pole count increases from 2 to 4 poles, for the same current and flux density. Considering the two rotors identical, with the same stack length, L_{rl} in (17.13) would not be much different. This leads us to the peak torque formula (17.3).

$$T_{ek} \approx \frac{3p_1}{2} \frac{(\Psi_{sph})^2}{2L_{sc}}; \Psi_{sph} = \frac{V_{ph}}{\omega_1} \quad (17.14)$$

For the same number of stator slots for $2p_1 = 2$ and 4, conductors per slot n_s , same airgap flux density, and same stator bore diameter, the phase flux linkage ratio in the two cases is

$$\frac{(\Psi_{sph})_{2p_1=2}}{(\Psi_{sph})_{2p_1=4}} = \frac{2}{1} \quad (17.15)$$

as the frequency is doubled for $2p_1 = 4$, in comparison with $2p_1 = 2$, for the same no load speed (f_1/p_1).

Consequently,

$$\frac{(L_{sc})_{2p_1=2}}{(L_{sc})_{2p_1=4}} = 1.5 - 1.8 \quad (17.16)$$

Thus, for the same bore diameter, stack length, and slot geometry,

$$\frac{(T_{ek})_{2p_1=2}}{(T_{ek})_{2p_1=4}} \approx \frac{2}{1.5 - 1.8} \quad (17.17)$$

From this simplified analysis, we may draw the conclusion that the $2p_1 = 2$ pole motor is better. If we consider the power factor and efficiency, we might end up with a notably better solution.

For super-high speed motors, $2p_1 = 2$ seems a good choice ($f_{1n} > 300$ Hz).

For a given total stator slot area (same current density and turns/phase), and the same stator bore diameter, increasing q_1 (number of slot/pole/phase) does not essentially influence L_{sl} (17.12) – the stator slot leakage and end connection leakage inductance components – as $n_s p_1 q_1 = W_1 = ct$ and slot depth remains constant while the slot width decreases to the extent q_1 increases, and so does λ_{end}

$$\lambda_{end} \approx \frac{0.34}{L_i} (l_{end} - 0.64 \cdot y) \cdot q_1 \quad (17.18)$$

with l_{end} = end connection length; y – coil throw; L_i – stack length.

However, λ_{ds} decreases and apparently so does λ_{zs} . In general, with larger q_1 , the total stator leakage inductance will decrease slightly. In addition, the stray losses have been proved to decrease with q_1 increasing.

A similar rationale is valid for the rotor leakage inductance L_{rl} (17.13) where the number of rotor slots increases. It is well understood that the condition $0.8N_s < N_r < N_s$ is to be observed.

A safe way to reduce the leakage reactance is to reduce the slot aspect ratio $h_{ss,r}/b_{ss,r} < 3.0-3.5$. For given current density this would lead to lower q_1 (or N_s) for a larger bore diameter, that is, a larger machine volume.

However, if the design current density is allowed to increase (sacrificing to some extent the efficiency) with a better cooling system, the slot aspect ratio could be kept low to reduce the leakage inductance L_{sc} .

A low leakage (transient) inductance L_{sc} is also required for current source inverter IM drives. [4]

So far, we have considered same current and flux densities, stator bore diameter, stack length, but the stator and yoke radial height for $2p_1 = 2$ is doubled with respect to the 4 pole machine.

$$\frac{(h_{cs,r})_{2p_1=2}}{(h_{cs,r})_{2p_1=4}} \approx \frac{(1.5-2)}{1} \quad (17.19)$$

Even if we oversaturated the stator and rotor yokes, and more for the two pole machine, the outer stator diameter will still be larger in the latter. It is true that this leads to a larger heat exchange area with the environment, but still the machine size is larger.

So, when the machine size is crucial, $2p_1 = 4$ [5] even $2p_1 = 6$ is chosen (urban transportation traction motors).

Whether to use long or short stack motors is another choice to make in the pursuit of smaller leakage inductance L_{sc} . Long stator stacks may allow smaller stator bore diameters, smaller pole pitches and thus smaller stator end connections.

Slightly smaller L_{sc} values are expected. However, a lower stator (rotor) diameter does imply deeper slots for the same current density. An increase in slot leakage occurs.

Finally, increasing the stack length leads to limited breakdown torque increase.

When low inertia is needed, the stack length is increased while the stator bore diameter is reduced. The efficiency will vary little, but the power factor will likely decrease. Consequently, the PEC KVA rating has to be slightly increased. The KVA ratings for two pole machines with the same external stator diameter and stack length, torque and speed, is smaller than for a 4 pole machine because of higher power factor. So when the inverter KVA is to be limited, the 2 pole machine might prevail.

A further way to decrease the stator leakage inductance may be to use four layers (instead of two) and chorded coils to produce some cancelling of mutual

leakage fluxes between them. The technological complication seems to render such approaches as less than practical.

17.7 WIDE CONSTANT POWER SPEED RANGE VIA VOLTAGE MANAGEMENT

Constant power speed range varies for many applications from 2 to 1 to 5(6) to 1 or more.

The obvious way to handle such requirements is to use an IM capable to produce, at base speed ω_b , a breakdown torque T_{bk} :

$$\frac{T_{bk}}{T_{en}} = \frac{\omega_{max}}{\omega_b} = C_\omega \quad (17.20)$$

- A larger motor

In general, IMs may not develop a peak to rated torque higher than 2.5 (3) to 1 (Figure 17.4a).

In the case when a large constant power speed range C_ω is required, it is only intuitive to use a larger IM (Figure 17.4b).

Adopting a larger motor, to have enough torque reserve up to maximum speed for constant power may be done either with an IM with $2p_l = 2,4$ of higher rating or a larger number of pole motor with the same power. While such a solution is obvious for wide constant power speed range ($C_\omega > 2.0 - 3.0$), it is not always acceptable as the machine size is notably increased.

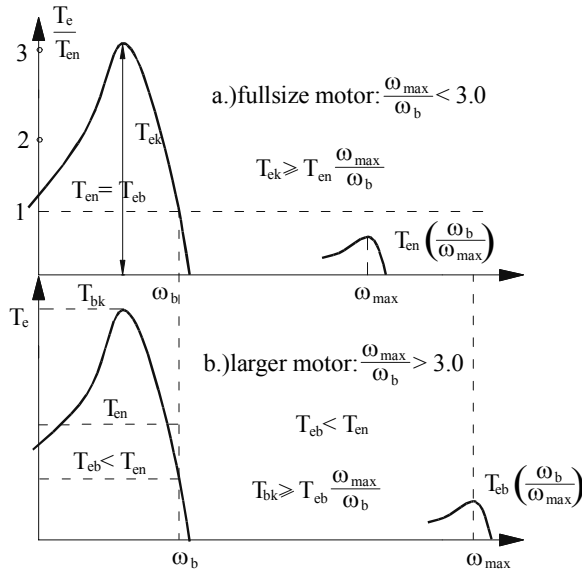


Figure 17.4 Torque-speed envelope for constant power
a) full size motor b) larger motor

- Higher voltage/phase

The typical torque/speed, voltage/speed, and current/speed envelopes for moderate constant power speed range are shown on [Figure 17.5](#).

The voltage is considered constant above base speed. The slip frequency f_{sr} is rather constant up to base speed and then increases up to maximum speed. Its maximum value $f_{sr\max}$ should be less or equal to the critical value (that corresponding to breakdown torque)

$$f_{sr} \leq f_{sr\max} = \frac{R_r}{2\pi L_{sc}} \quad (17.21)$$

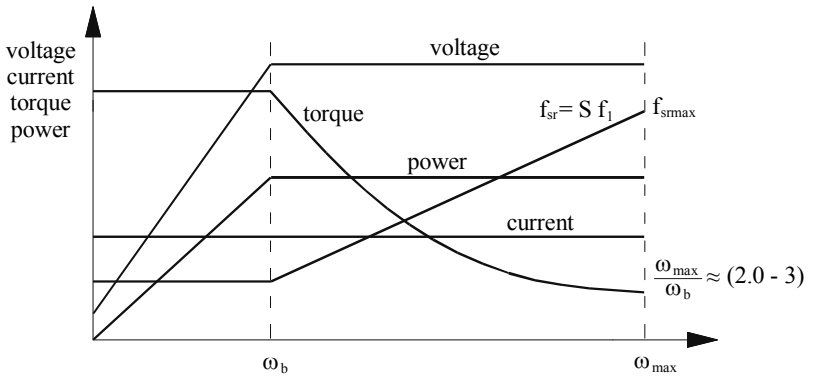


Figure 17.5 Torque, voltage, current versus frequency

$$(T_{bk})_{\omega_{\max}} = \frac{3}{2} p_l \left(\frac{V_{ph}}{\omega_{r\max}} \right)^2 \cdot \frac{1}{L_{sc}} \geq T_{eb} \cdot \frac{\omega_b}{\omega_{\max}} \quad (17.22)$$

If

$$(T_{bk})_{\omega_{\max}} < T_{eb} \frac{\omega_b}{\omega_{\max}}, \quad (17.23)$$

it means that the motor has to be oversized to produce enough torque at maximum speed.

If a torque (power reserve) is to be secured for fast transients, a larger torque motor is required.

Alternatively, the phase voltage may be increased during the entire constant power range ([Figure 17.6](#))

To provide a certain constant overloading over the entire speed range, the phase voltage has to increase over the entire constant power speed range. This

means that for base speed ω_b , the motor will be designed for lower voltage and thus larger current. The inverter current loading (and costs) will be increased.

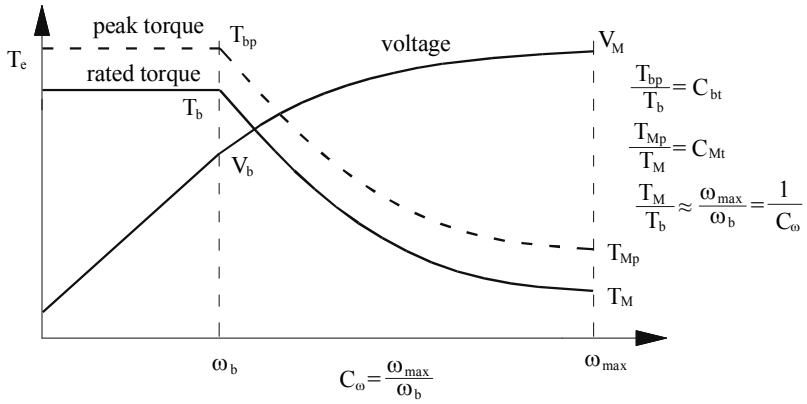


Figure 17.6 Raising voltage for the constant power speed range

Let us calculate the torque T_e , and stator current I_s from the basic equivalent circuit.

$$T_e \approx \frac{3p_l}{\omega_l} \cdot \frac{V_{ph}^2 \cdot R_r / s}{\left(R_s + C_l \frac{R_r}{s} \right)^2 + \omega_l^2 L_{sc}'^2}; \quad (17.24)$$

$$L_{sc}' = L_{sl} + C_l L_{rl}; C_l = 1 + \frac{L_{sl}}{L_m} \approx 1.02 - 1.08$$

$$I_s \approx V_{ph} \sqrt{\left(\frac{1}{\omega_l L_m} \right)^2 + \frac{1}{\left(R_s + C_l \frac{R_r}{s} \right)^2 + \omega_l^2 L_{sc}'^2}} \quad (17.25)$$

At maximum speed, the machine has to develop the peak torque T_{Mp} at maximum voltage V_M .

Now if for both T_{bp} and T_{Mp} , the breakdown torque conditions are met,

$$T_{bp} = \frac{3p_l}{2} \left(\frac{V_b}{\omega_b} \right)^2 \cdot \frac{1}{2L_{sc}}, \quad (17.26)$$

$$T_{Mp} = \frac{3p_l}{2} \left(\frac{V_M}{\omega_{max}} \right)^2 \cdot \frac{1}{2L_{sc}},$$

the voltage ratios V_M/V_b is (as in [6])

$$\left(\frac{V_M}{V_b}\right)^2 = \left(\frac{\omega_{\max}}{\omega_b}\right)^2 \frac{T_{Mp}}{T_{bp}} = \frac{\omega_{\max}}{\omega_b} \cdot \frac{\left(\frac{T_{Mp}}{T_M}\right)}{\left(\frac{T_{bp}}{T_b}\right)} = C_\omega \frac{C_{Mt}}{C_{bt}} \quad (17.27)$$

For $\omega_{\max}/\omega_b = 4$, $C_{Mt} = 1.5$, $C_{bt} = 2.5$, we find from (17.17) that $V_M/V_b = 1.55$. So the current rating of the inverter (motor) has to be increased by 55%.

Such a solution looks extravagant in terms of converter extra-costs, but it does not suppose IM overrating (in terms of power). Only a special design for lower voltage V_b at base speed ω_b is required.

For

$$S = S_{K,b,m} = \frac{C_1 R_r}{\sqrt{R_s^2 + \omega_{b,M}^2 L_{sc}^2}} \quad (17.28)$$

and $\omega_1 = \omega_b$ and, respectively, $\omega_1 = \omega_M$ from (17.25), the corresponding base and peak current values I_b , I_M may be obtained.

These currents are to be used in the motor (I_b) and converter ($I_M \sqrt{2}$) design.

Now, based on the above rationale, the various cases, from constant overloading $C_{MT} = C_{bt}$ to zero overloading at maximum speed $C_{MT} = 1$, may be treated in terms of finding V_M/V_b and I_M/I_b ratios and thus prepare for motor and converter design (or selection). [6]

As already mentioned, reducing the slot leakage inductance components through lower slot aspect ratios may help to increase the peak torque by as much as (50–60%) in some cases, when forced air cooling with a constant speed fan is used. [6]

- High constant power speed range: $\omega_{\max}/\omega_b > 4$

For a large constant power speed range, the above methods are not generally sufficient. Changing the phase voltage by switching motor phase connection from star to delta (Y to Δ) leads to a sudden increase in phase voltage by $\sqrt{3}$. A notably larger constant power speed range is obtained this way (Figure 17.7).

The Y to Δ connection switching has to be done through a magnetic switch with the PEC control temporarily inhibited and then so smoothly reconnected that no notable torque transients occurred. This voltage change is done “on the fly.”

An even larger constant power speed range extension may be obtained by using a winding tap (Figure 17.8) to switch from a large to a small number of turns per phase (from W_1 to W'_1 , $W_1/W'_1 = C_{W1} > 1$).

A double single-throw magnetic switch suffices to switch from high (W_1) to low number of turns (W_1/C_{W1}) for high speed.

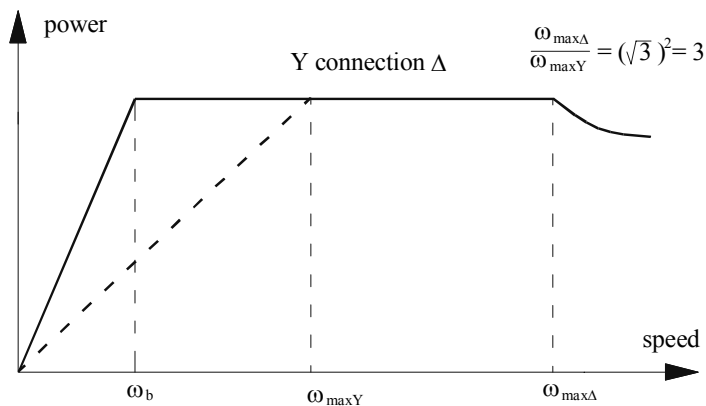


Figure 17.7 Extended constant power speed range by Y to Δ stator connection switching

Once the winding is designed with the same current density for both winding sections, the switching from W_1 to W_1/C_{W1} turns leads to stator and rotor resistance reduction by $C_{W1} > 1$ times and a total leakage inductance L_{sc} reduction by C_{W1}^2 .

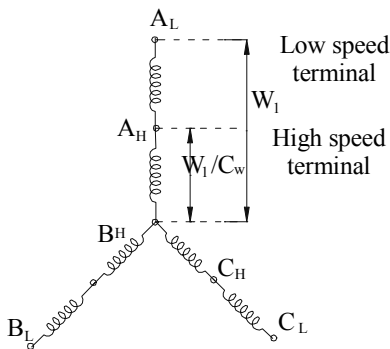


Figure 17.8 Tapped stator winding

In terms of peak (breakdown torque), it means a strong increase by C_{W1}^2 . The constant power speed range is extended by

$$\frac{\omega_{\max H}}{\omega_{\max L}} = C_{W1}^2 \quad (17.29)$$

The result is similar to that obtained with Y to Δ connection switching but this time C_{W1} may be made larger than $\sqrt{3}$ and thus a larger extension of

constant power speed range is obtained without oversizing the motor. The price to be paid is the magnetic (or thyristor-soft) switch. The current ratios for the two winding situations for the peak torque are:

$$\frac{I_{s\max H}}{I_{s\max L}} \approx C_{W1} \quad (17.30)$$

The leakage inductance L_{sc} decreases with C_W^2 while the frequency increases C_W times. Consequently, the impedance decreases, at peak torque conditions, C_W times. This is why the maximum current increases C_{W1} times at peak speed $\omega_{\max H}$ with respect to its value at $\omega_{\max L}$. Again, the inverter rating has to be increased accordingly while the motor cooling must be adequate for these high demands in winding losses. The core losses are much smaller but they are generally smaller than copper losses unless super-high speed (or large power) is considered. As for how to build the stator winding, to remain symmetric in the high speed connection, it seems feasible to use two unequal coils per layer and make two windings, with W_1/C_{W1} , and $W_1(1 - C_{W1}^{-1})$ turns per phase, respectively, and connect them in series. The slot filling factor will be slightly reduced but, with the lower turns coil on top of each slot layer, a further reduction of slot leakage inductance for high speed connection is obtained as a bonus.

• Inverter pole switching

It is basically possible to reduce the number of poles in the ratio of 2 to 1 through using two twin half-rating inverters and thus increase the constant power speed range by a factor of two. [7]

No important oversizing of the converter seems needed. In applications when PEC redundancy for better feasibility is a must, such a solution may prove adequate.

It is also possible to use a single PEC and two simple three phase soft switches (with thyristors) to connect it to the two three phase terminals corresponding to two different pole count windings ([Figure 17.9](#))

Such windings with good winding factors are presented in Chapter 4. As can be seen from [Figure 17.9](#), the current remains rather constant over the extended constant power speed range, as does the converter rating. Apart from the two magnet (or static) switches, only a modest additional motor over-cost due to dual pole count winding is to be considered the price to pay for speed range extension at constant power. For cost sensitive applications, this solution might be practical.

17.8 DESIGN FOR HIGH AND SUPER-HIGH SPEED APPLICATIONS

As previously mentioned, super-high speeds range start at about 18 krpm (300 Hz in 2 pole IMs). Between 3 krpm (50 Hz) and 18 krpm (300 Hz) the interval of high speeds is located ([Figure 17.10](#))

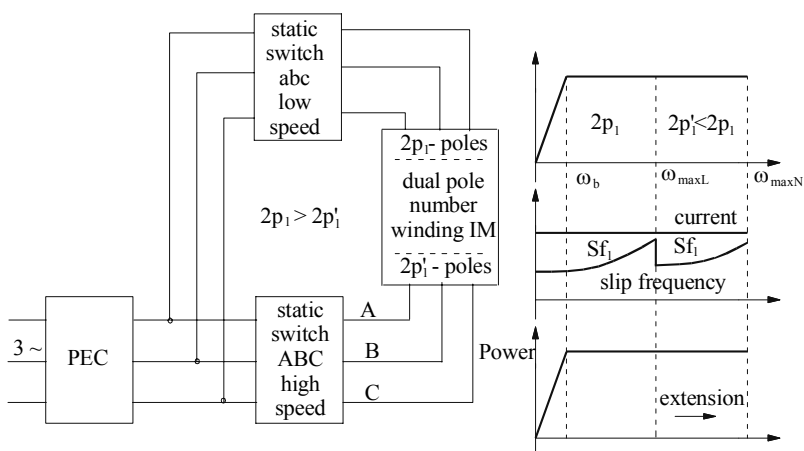


Figure 17.9 Dual static switch for dual pole count winding IM

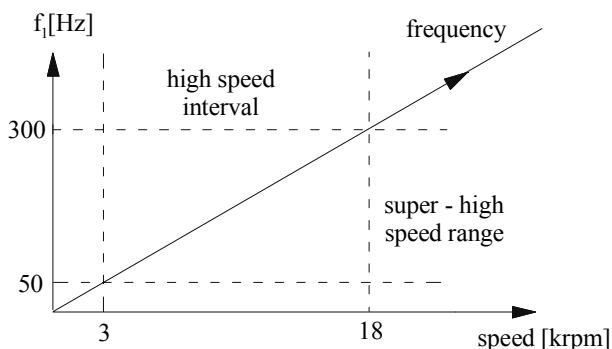


Figure 17.10 High and super-high speed division

Below 300 Hz, where standard PWM-PEC are available, up to tens of kW (recently up to MW/unit) but above 50 Hz, the design is very much similar to that for speeds below 50 Hz. Two pole motors are favored, with lower flux densities to limit core losses, forced stator cooling, and carefully limited mechanical losses. Laminations and conductors remain of the standard type with close watch on skin effect containment.

Above 300 Hz, in the super-high speed applications, the increase in mechanical stresses, mechanical and core loss, and in skin effects have triggered specialized worldwide research efforts. We will characterize the most representative of them in what follows.

17.8.1 Electromagnetic limitations

The Esson's constant C_0 still holds (Chapter 14, Equation (14.14)),

$$C_0 \approx \frac{\pi^2}{\sqrt{2}} K_{w1} A_1 B_g \quad (17.31)$$

The linear current density A_1 (Aturns/m) is limited to $(5 - 20) \times 10^3$ A/m for most super-high speed IMs, and in general increases with stator bore diameter D_{is} .

The stator bore diameter D_{is} is related to motor power P_n as (Chapter 14, Equation (14.15))

$$D_{is} = \sqrt{\frac{K_E P_n}{\eta \cos \phi} \cdot \frac{p_1}{f_1} \cdot \frac{1}{C_0} \cdot \frac{1}{L}}; K_E \approx 0.98 - p_1 \cdot 5 \cdot 10^{-3} \quad (17.32)$$

f_1 – frequency, p_1 – pole pairs, L – stack length, η – efficiency, K_E – e.m.f. coefficient.

As the no load speed $n_1 = f_1/p$ is large, for given stack length L , the stator bore diameter D_{is} may be obtained for given C_0 , from (7.32).

For super-high speed motors, even with thin laminations (0.1 mm thick or less), the airgap flux density is lowered to $B_g < (0.5 - 0.6)T$ to cut the core losses which tend to be large as frequency increases. Longer motors (L – larger) tend to require, as expected, low stator bore diameters D_{is} .

17.8.2 Rotor cooling limitations

Due to the uni-stack rotor construction, made either from solid iron or from laminations, the heat due to rotor losses – total conductor, core and air friction losses – has to be transferred to the cooling agent almost entirely through the rotor surface. Thus, the rotor diameter $D_{er} \approx D_{is}$ is limited by

$$D_{is} \geq \frac{\text{rotor losses}}{\pi L \cdot \alpha \Delta \theta_{ra}} \quad (17.33)$$

α – heat transfer factor $W/m^2 \text{ } ^0K$, $\Delta \theta_{ra}$ – rotor to internal air temperature differential.

Among rotor losses, besides fundamental cage and core losses, space and time harmonics losses in the core and rotor cage are notable.

17.8.3 Rotor mechanical strength

The centrifugal stress on the rotor increases with speed squared. At the laminated rotor bore $D_{er} \approx D_{is}$, the critical stress $\sigma_K (N/m^2)$ should not be surpassed. [6]

$$D_{is} < \sqrt{\frac{16\sigma_K}{\gamma \omega_r^2 (3 + \mu)} - \frac{(1 - \mu)}{3 + \mu}} \cdot D_{int}^2 \quad (17.34)$$

γ – rotor material density (kg/m^3), ω_r – rotor angular speed (rad/s), μ – Poisson's ratio, $D_{int}(m)$ – interior diameter of rotor laminations ($D_{int} = 0$ for solid rotor).

For 100 kW IMs with $L/D_{is} = 3$ (long stack motors) and $C_0 = 60 \times 10^3 \text{ J/m}^3$, the limiting curves given by (17.32) – (17.34) are shown in Figure 17.12. [8]

Results in Figure 17.11 lead to remarks such that

- To increase the speed, improved (eventually liquid) rotor cooling may be required.
- When speed and the Esson's constant increase, the rotor losses per rotor volume increase and thus thermal limitations become the main problem.
- The centrifugal stress in the laminated rotor restricts the speed range (for 100 kW).

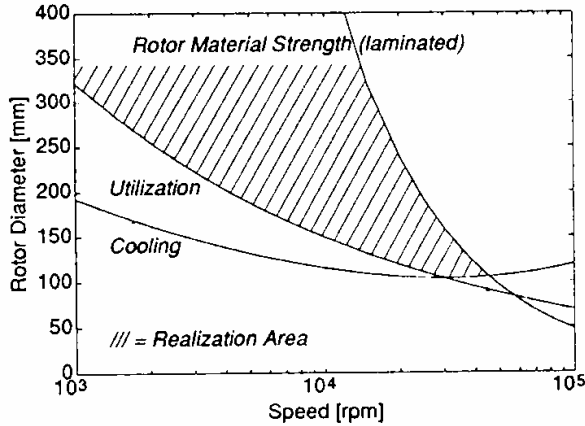


Figure 17.11 Stator bore diameter limiting curves for 100 kW machines at very high speed

17.8.4 The solid iron rotor

As the laminated rotor shows marked limitations, extending the speed range for given power leads, inevitably, to the solid rotor configuration. The absence of the central hole and the solid structure produce a more rugged configuration.

The solid iron rotor is also better in terms of heat transmission, as it allows for good axial heat exchange by thermal conductivity.

Unfortunately, the smooth solid iron rotor is characterized by a large equivalent resistance R_s and a rather limited magnetization inductance. This is so because the depth of field penetration in the rotor δ_{iron} is [9]

$$\delta_i \approx \left| \operatorname{Re} \left(\sqrt{\frac{1}{(\pi/\tau)^2 + j2\pi f_i S \frac{\sigma_{iron}}{K_T} \cdot \mu_0 \cdot \mu_{rel}}} \right) \right| \quad (17.35)$$

τ – pole pitch, μ_{rel} – relative iron permeability, K_T – conductivity correction factor to be explained later in this paragraph:

$$\begin{aligned} a &= (\pi / \tau)^2 \\ b &= 2\pi f_1 S \frac{\sigma_{\text{iron}}}{K_T} \cdot \mu_0 \mu_{\text{rel}} \end{aligned} \quad (17.36)$$

$$\delta_i = \frac{\sqrt{a + \sqrt{a^2 + b^2}}}{\sqrt{2(a^2 + b^2)}} \quad (17.37)$$

For $\tau = 0.1\text{m}$, $\mu_{\text{rel}} = 18$, $S_{f_1} = 5\text{ Hz}$, $\sigma_{\text{iron}} = 10^{-1}\sigma_{\text{co}} = 5 \times 10^6 (\Omega\text{m})^{-1}$, from (17.36 – 17.37), $\delta_i = 1.316 \times 10^{-2}\text{m}$.

To provide reasonable airgap flux density $B_g \approx 0.5 - 0.6\text{T}$ the field penetration depth in iron δ_i for a highly saturated iron, with $\mu_{\text{rel}} = 36$ (for $S_{f_1} = 5\text{ Hz}$) and $B_{\text{iron}} = 2.3\text{T}$ ($H_{\text{iron}} = 101733.6\text{ A/m}$), must be

$$\delta_{\text{iron}} \geq \frac{B_g \cdot \tau / \pi}{B_{\text{iron}}} = \frac{0.55}{\pi \times 2.3} \times \tau = 7.6 \cdot 10^{-2} \times \tau \text{ [m]} \quad (17.38)$$

For $\delta_{\text{iron}} = 0.01316\text{ m}$, the pole pitch τ should be $\tau < 0.173\text{ m}$. As the pole pitch was supposed to be $\tau = 0.1\text{m}$, the iron permeability may be higher.

For a $2p_1 = 2$ pole machine, typical for super-high speed IMs, the stator bore diameter D_{is} ,

$$D_{\text{is}} < \frac{2p_1 \tau}{\pi} = \frac{2 \times 1 \times 0.173}{\pi} = 0.110\text{ m!} \quad (17.39)$$

This is a severe limitation in terms of stator bore diameter (see [Figure 17.12](#)). For higher diameters D_{is} , smaller relative permeabilities have to be allowed for. The consequence of the heavily saturated rotor iron is a large magnetization m.m.f. contribution which in relative values to airgap m.m.f. is

$$\frac{F'_{\text{iron}}}{F_{\text{airgap}}} \approx \frac{\frac{1}{3} \tau \times H_{\text{iron}} \cdot \mu_0}{B_g \cdot 2g} \approx \frac{\frac{1}{3} 0.1 \times 101733 \times 1.256 \times 10^{-6}}{0.55 \times 2 \times 1.0 \times 10^{-3}} = 3.872! \quad (17.40)$$

A rather low power factor is expected. The rotor leakage reactance to resistance ratio may be approximately considered as 1.0 (for constant, even low permeability)

$$\frac{\omega_1 L_{rl}}{R_r} \approx 1 \quad (17.41)$$

From [8] the rotor resistance reduced to the stator is

$$R_r = \frac{6L \cdot K_T}{\delta_i \tau p_1 \sigma_{\text{iron}}} \cdot (K_{w1} W_1)^2 \quad (17.42)$$

In (17.42), the rotor length was considered equal to stator stack length and K_T is a transverse edge coefficient which accounts for the fact that the rotor current paths have a circumferential component. (Figure 17.12)

From [8],

$$K_T \approx \frac{1}{1 - \frac{\tanh(\pi L / 2\tau \cdot n)}{(\pi L / 2\tau \cdot K)}} > 1 \quad (17.43)$$

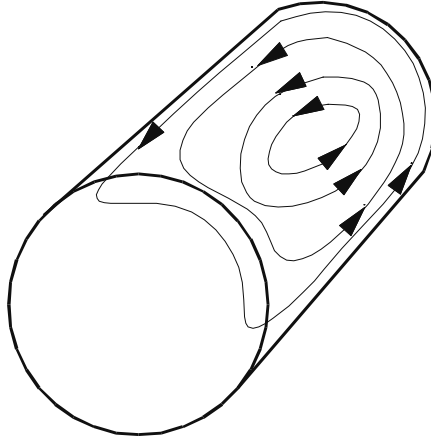


Figure 17.12 Smooth solid rotor induced current paths

The longer the stack length/pole pitch ratio L/τ , the smaller is this coefficient.

Here $n = 1$. While K_T reduces the equivalent apparent conductivity of iron, it does increase the rotor resistivity as $\sqrt{K_T}$, lowering the torque for given slip frequency Sf_1 .

This is why the rotor structure may be slitted with the rotor longer than the stack to get lower transverse coefficient. With copper end rings, we may consider $K_T = 1$ in (17.43) and in (17.35), (since the end rings resistance is much smaller than the rotor iron resistance (Figure 17.13)).

How deep the rotor slits should be is a matter of optimal design as the main flux paths have to close below their bottom. So definitely $h_{\text{slit}} > \delta_i$. Notably larger output has been demonstrated for given stator with slitted rotor and copper end rings. [10]

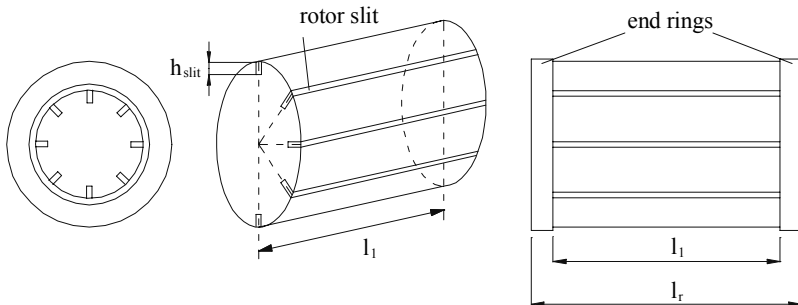


Figure 17.13 Radially slitted rotor with end rings

The radial slots may be made wider and filled with copper bars (Figure 17.14a); copper end rings are added. As expected, still larger output for better efficiency and power factor has been obtained. [10]

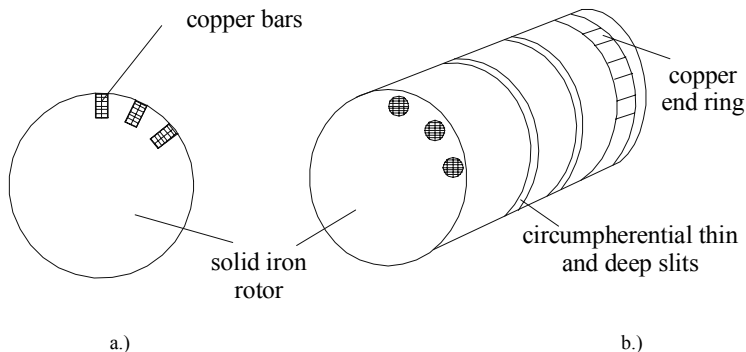


Figure 17.14 Solid iron rotor with copper bars

a.) copper bars in radial open slots (slits)

b.) copper bars in closed slots

Another solution would be closed rotor slots with copper bars and circumferential deep slits in the rotor (Figure 17.14b).

This time, before the copper bars are inserted in slots the rotor is provided with thin and deep circumferential slits such that to increase the transverse edge effect coefficient K_T (17.43) where n is the number of stack axial sections produced by slits.

This time they serve to destroy the solid iron eddy currents as the copper bars produce more efficiently torque, that is for lower rotor resistance. The value of n may be determined such that K_T is so large that depth of penetration $\delta_{iron} \geq 2d_{co}$ (d_{co} – copper bar diameter). Three to six such circumferential slits generally suffice.

The closed rotor slots serve to keep the copper bar tight while the slitted iron serves as a rugged rotor and better core and the space harmonics losses in the rotor iron are kept reasonably small.

A thorough analysis of the solid rotor with or without copper cage with and without radial or circumferential slits may require a full fledged 3D FEM approach. This is time consuming and thus full scale careful testing may not be avoided.

17.8.5 21 kW, 47,000 rpm, 94% efficiency with laminated rotor [11]

As seen in [Figure 17.12](#), at 50,000 rpm, a stator bore diameter D_{is} of up to 80 mm is acceptable from the rotor material strength point of view.

Reference [5] reports a good performance 21 kW, 47,000 rpm ($f_1 = 800$ Hz) IM with a laminated rotor which is very carefully designed and tested for mechanical and thermal compliance.

High silicon 3.1% laminations (0.36 mm thick) were specially thermally treated to reduce core loss at 800 Hz to less than 30 W/kg at 1T, and to avoid brittleness, while still holding a yield strength above 500 MPa. Closed rotor slots are used and the rotor lamination stress in the slot bridge zone was carefully investigated by FEM.

Al 25 (85% copper conductivity) was used for the rotor bars, while the more rugged Al 60 (75% copper conductivity) was used to make the end rings. The airgap was $g = 1.27$ mm for a rotor diameter of 51 mm, and a stack length $L = 102$ mm and $2p_1 = 2$ poles. The stator/rotor slot numbers are $N_s/N_r = 24/17$, rated voltage 420 V, and current 40 A. A rather thick shaft is provided (30 mm in diameter) for mechanical reasons. However, in this case part of it is used as rotor yoke for the main flux. As the rated slip frequency $Sf_1 = 5$ Hz the flux penetrates enough in the shaft to make the latter a good part of the rotor yoke. The motor is used as a direct drive for a high speed centrifugal compressor system. Shaft balancing is essential. Cooling of the motor is done in the stator by using a liquid cold refrigerant. The refrigerant also flows through the end connections and through the airgap zone in the gaseous form. To cut the pressure drop, a rather large airgap ($g = 1.27$ mm) was adopted. The motor is fed from an IGBT voltage source PWM converter switched at 15 kHz.

The solution proves to be the least expensive by a notable margin when compared with PM-brushless and switched reluctance motor drives of equivalent performance.

This is a clear example how, by carefully pushing forward existing technologies new boundaries are reached. [11,12]

17.9 SAMPLE DESIGN APPROACH FOR WIDE CONSTANT POWER SPEED RANGE

Design specifications are

Base power: $P_n = 7.5$ kW

Base speed: $n_b = 500$ rpm

Maximum speed $n_{max} = 6000$ rpm

Constant power speed range: 500 rpm to 6000 rpm

Power supply voltage: 400 V, 50 Hz, 3 phase

Solution characterization

The design specifications indicate an unusually large constant power speed range $n_{\max}/n_b = 12:1$. Among the solutions for this case, it seems that a combination of winding connection switching from Y to Δ and high peak torque design with lower than rated voltage at base speed might do it.

The Y to Δ connection produces an increase of constant power speed zone by

$$\frac{n_{\max\Delta}}{n_{\max Y}} = \left[\frac{(V_{ph})_{\Delta}}{(V_{ph})_Y} \right]^2 = \frac{3}{1} \quad (17.44)$$

It remains a 4/1 ratio constant power speed range for the Y (low speed) connection of windings.

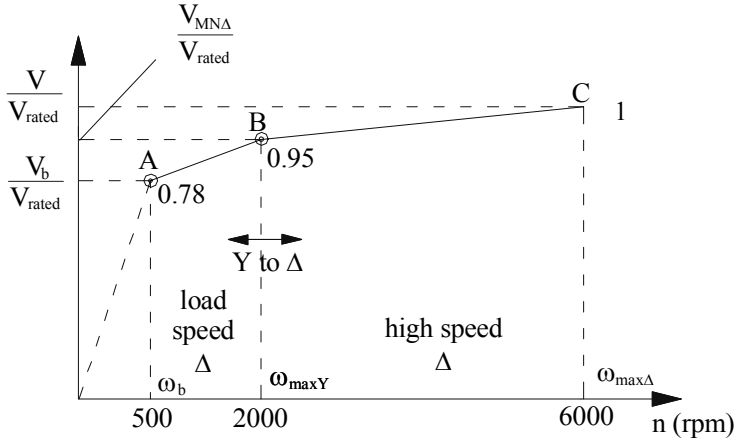


Figure 17.15 Line voltage envelope versus speed

As the peak/rated torque seldom goes above 2.5-2.7, the voltage at base speed has to be reduced notably.

To leave some torque reserve up to maximum speed, even in the Δ (high speed) connection, a voltage reserve has to be left.

First, from (17.27), we may calculate the $V_{MY/\Delta}/V_{rated}$ from the high speed zone,

$$\left(\frac{V_{MY/\Delta}}{V_{rated}} \right)^{-2} = \frac{\omega_{\max\Delta}}{\omega_{\max Y}} \cdot \frac{C_{MT}}{C_{bT}} \quad (17.45)$$

C_{MT} is the overload capacity at maximum speed. Let us consider (C_{MT}) = 1 for $\omega_{\max\Delta}$ (6000 rpm). C_{bT} is the ratio between peak torque and base torque, a matter of motor design. Let us take $C_{bT} = 2.7$. In this case from (17.45)(point B),

$$\left(\frac{V_{MY/\Delta}}{V_{rated}} \right) = \sqrt{\frac{2.7}{3 \times 1}} = 0.95 \quad (17.46)$$

Now for point A in [Figure 17.15](#), we may apply the same formula,

$$\left(\frac{V_b}{V_{rated}} \right)^{-2} = \left(\frac{V_{mY/\Delta}}{V_{rated}} \right)^{-2} \cdot \frac{\omega_{maxY}}{\omega_b} \cdot \frac{C_{Mt}}{C_{bt}} \quad (17.46')$$

As for point A $V < V_{rated}$, there is a torque reserve, so we may consider again $C_{Mt} = 1.0$

$$\frac{V_b}{V_{rated}} = 0.95 \sqrt{\frac{1}{4} \cdot \frac{2.7}{1.0}} = 0.78 \quad (17.47)$$

Admitting a voltage derating of 5% ($V_{derat} = 0.05$) due to the PEC, the base line voltage at motor terminals, which should produce rated power at $n_b = 500$ rpm, is

$$(V_b)_{line} = V_{line} \cdot \frac{V_b}{V_{rated}} \cdot (1 - V_{derat}) = 400 \times 0.78(1 - 0.05) = 296V \quad (17.48)$$

For constant power factor and efficiency, reducing the voltage for a motor does increase the current in proportion:

$$\frac{I'_b}{I_b} \approx \frac{V_{rated}}{V_b} = \frac{400}{296} = 1.35 \quad (17.49)$$

As this overloading is not needed during steady state, it follows that the power electronics converter does not need to be overrated as it allows 50% current overload for short duration by design.

Now the largest torque T_b is needed at base speed (500 rpm), with lowest voltage $V_{bline} = 296$ V.

$$T_b = \frac{P_b}{2\pi \cdot \frac{n_b}{60}} = \frac{7500 \times 60}{2\pi \cdot 500} = 143.3Nm \quad (17.50)$$

The needed peak (breakdown) torque T_{bk} is

$$T_{bk} = C_{bT} \times T_b = 2.7 \times 143.3 = 387Nm \quad (17.51)$$

The electromagnetic design has to be done for base power, base speed, base voltage with the peak torque ratio constraint.

The thermal analysis has to consider the operation times at different speeds. Also the core (fundamental and additional) and mechanical losses increase with speed, while the winding losses also slightly increase due to power factor decreasing with speed and due to additional winding loss increase with frequency.

Basically, from now on the electromagnetic design is similar to that with constant V/f (Chapter 15,16). However, care must be exercised to additional losses due to skin effects, space and time harmonics, and their influence on temperature rise and efficiency.

17.10 SUMMARY

- Variable speed drives with induction motors may be classified by the speed range for constant output power.
- Servodrives lack any constant power speed range and the base torque is available for fast acceleration and deceleration provided the inverter is oversized to handle the higher currents typical for breakdown torque conditions.
- General drives have a moderate constant power speed range of $\omega_{\max}/\omega_b < 2.5$, for constant (ceiling) voltage from the PEC. Increases in the ratio $\omega_{\max}/\omega_b$ are obtainable basically only by reducing the total leakage inductance of the machine. Various parameters such as stack length, bore diameter, or slot aspect ratio may be used to increase the “natural” constant power speed range up to $\omega_{\max}/\omega_b \leq T_{bk}/T_b < (2.2 - 2.7)$. No overrating of the PEC or of the IM is required.
- Constant power drives are characterized by $\omega_{\max}/\omega_b > T_{bk}/T_{eb}$. There are quite a few ways to get around this challenging task. Among them is designing the motor at base speed for a lower voltage and allowing the voltage to increase steadily and reach its peak value at maximum speed.
- The constant power speed zone may be extended notably this way but at the price of higher rated (base) current in the motor and PEC. While the motor does not have to be oversized, the PEC does. This method must be used with cautious.
- Using a winding tap, and some static (or magnetic) switch, is another very efficient approach to extend the constant power speed zone because the leakage inductance is reduced by the number of turns reduction ratio for high speeds. A much higher peak torque is obtained. This explains the widening of constant power speed range. The current raise is proportional to the turns reduction ratio. So PEC oversizing is inevitable. Also, only a part of the winding is used at high speeds.
- Winding connection switching from Y to Δ is capable of increasing the speed range at constant power by as much as 3 times without overrating the motor or the converter. A dedicated magnetic (or static) switch with PEC short time shutdown is required during the connection switching.
- Pole changing from higher to lower also extends the constant power speed zone by p_1/p_2 times. For $p_1/p_2 = 2/1$, a twin half rated power PEC may be used to do the switching of pole count. Again, no oversizing of motor or total converter power is required. Alternatively, a single full power PEC may be used to supply a motor with dual pole

count winding terminals by using two conventional static (or magnetic) three phase power switches.

- The design for variable speed has to reduce the skin effect which has a strong impact on rotor cage sizing.
- The starting torque and current constraints typical for constant V/f design do no longer exist.
- High efficiency IMs with insulated bearings designed for constant V/f may be (in general) safely supplied by PEC to produce variable speed drives.
- Super-high speed drives (centrifugal compressors, vacuum pumps, spindles) are considered here for speeds above 18,000 rpm (or 300 Hz fundamental frequency). This classification reflects the actual standard PWM-PECs performance limitations
- For super-high speeds, the torque density is reduced and thus the rotor thermal and mechanical limitations become even more evident. Centrifugal force – produced mechanical stress might impose, at certain peripheral rotor speed, the usage of a solid iron rotor.
- While careful design of laminated rotor IMs has reached 21 kW at 47,000 rpm (800 Hz), for larger powers and equivalent or higher peripheral speed, solid iron rotors may be mandatory.
- To reduce iron equivalent conductivity and thus increase field penetration depth, the solid iron rotor with circumferential deep (and thin) slits and copper bars in slots seems a practical solution.
- The most challenging design requirements occur for wide constant power speed range applications ($\omega_{\max}/\omega_b > \text{peak_torque/base_torque}$). The electromagnetic design is performed for base speed and base power. The thermal design has to consider the loss evolution with increasing speed.
- Once the design specifications and the most demanding performance are identified, the design (sizing) of IM for variable speed may follow a similar path as for constant V/f operation (Chapter 14–16).

17.11 REFERENCES

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